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# Design Considerations for Nonlinear Scattering

by Christian Fazi, Frank Crowne, and Marc Ressler

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Renewed interest in harmonic and multitone radar has motivated design efforts based on the standard approach found in the radar literature, i.e., the classical radar equation. In this report, we show that such approaches can be problematic in determining the correct power budget requirements for detecting a nonlinear target, and that it is necessary to examine the differences of a linear versus nonlinear target cross-section.

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# 1. Introduction

There has been recent interest in the use of harmonic and multitone radar for various applications, from clutter suppression during tracking of animals and insects (*1*), to improving target classification when detecting radar tags (*2*). In designing systems based on these new technologies, it is incumbent on the designer to resist the temptation to use traditional methodologies carelessly. This applies in particular to the radar equation, which must be modified so as to incorporate the way a nonlinear scatterer (target) responds to electromagnetic (EM) illumination, which (as we will see further) differs significantly from the response of an ordinary target.

# 2. Linear Radar

In order to discuss nonlinear effects in radar, we must first specify what is meant by a *linear target*. It is reasonable to identify a target as linear when, upon illumination by an EM field, it radiates a scattered field that is directly proportional to the incident field. For a monostatic radar, this makes it possible to reduce the properties of the scatterer to a single parameter, the (backscattering) *cross-section* (*3*), which is purely a property of the scatterer and independent of any property of the excitation source. This definition leads to the well-established radar equation (*4*). Assuming such a linear target is illuminated by an antenna with area $A$ at a distance $R$, the antenna gain is:

$$G = \frac{4\pi A\rho}{\lambda^2},$$

where $\lambda$ is the radar wavelength and ρ is the antenna efficiency, which we will assume to be 1. A power $P$ fed to this antenna creates a power *flux* $\wp$ (watts/area) at the target of

$$\wp = \frac{PG}{4\pi R^2},$$

and, hence, an electric field at the target

$$E_{inc} = (Z\wp)^{1/2} = \frac{(ZPG)^{1/2}}{(4\pi)^{1/2} R}.$$

Here, $Z$ is the antenna's impedance. At the target, this field induces a current, which in turn radiates a scattered field back to the antenna:

$$E_{sc} = \frac{\sigma^{1/2}}{R} E_{inc} = \frac{(\sigma ZPG)^{1/2}}{(4\pi)^{1/2} R^2},$$

where $\sigma$ is the radar cross-section. If the distance $R$ back to the receiver is large, this scattered field consists of a plane wave whose amplitude has fallen off by $1/R^2$. Then, once more taking into account the receiver gain, the field induced at the receiver is

$$E_{rec} = \left(\frac{G}{4\pi}\right)^{1/2} \cdot E_{sc} = \frac{(\sigma ZP)^{1/2} G}{4\pi R^2},$$

and the received power is

$$\begin{aligned} P_{rec} &= Z^{-1} E_{rec}{}^2 A = \sigma \frac{PGA}{(4\pi)^2 R^4} \\ &= \sigma \frac{PG^2 \lambda^2}{(4\pi)^3 R^4} \end{aligned},$$

which is proportional to $1/R^4$, the classical result.

## 3. Nonlinear Radar

This derivation fails for a nonlinear scatterer. To illustrate why, let us assume that the nonlinearity of the scatterer is small enough to admit a power-law expansion, a situation that illustrates our argument and is also simple mathematically (although other associated types of nonlinearities—e.g., clipping, Schottky barriers, etc.—lead to similar results). Let us illuminate the target with two signals, $E_1$ and $E_2$, close enough in frequency that the antenna can emit and receive both efficiently. We then assume that the response of the scatterer takes the form

$$I_{sc} = \sigma^{1/2} E_{inc} + \alpha E_{inc}{}^2 + \beta E_{inc}{}^3,$$

where $\alpha$ and $\beta$ are constants that are specific to the target, and $I_{sc}$ is the current at the target generated by the incident field $E_{inc}$. Since this field contains two fields, $E_1$ and $E_2$, at two frequencies, $\omega_1$ and $\omega_2$, the current $I_{sc}$ will also contain components at these frequencies. For a linear target, there will be no frequencies in the return signal other than these. However, if we view the target as a ~~"d~~evice" in the circuit-theoretic sense, we know that the current induced by $E_{inc}$ must also contain additional frequencies. Thus, if

$$E_{inc} = E_1 \cos(\omega_1 t + \phi_1) + E_2 \cos(\omega_2 t + \phi_2),$$

the contributions to $I_{sc}$ from the squared term can be written

$$
\begin{aligned}
&E_{inc}{}^{2} = \left[E_1 \cos\left(\omega_1 t + \phi_1\right) + E_2 \cos\left(\omega_2 t + \phi_2\right)\right]^2 \\
&= E_1{}^{2} \cos\left(\omega_1 t + \phi_1\right)^2 + 2E_1 E_2 \cos\left(\omega_1 t + \phi_1\right)\cos\left(\omega_2 t + \phi_2\right) \\
&+ E_2{}^{2} \cos\left(\omega_2 t + \phi_2\right)^2 \\
&= \frac{1}{2}E_1{}^{2}\left[1 + \cos\left(2\omega_1 t + 2\phi_1\right)\right] + \frac{1}{2}E_2{}^{2}\left[1 + \cos\left(2\omega_2 t + 2\phi_2\right)\right] \\
&+ E_1 E_2\left[\cos\left(\left[\omega_1 - \omega_2\right]t + \phi_1 - \phi_2\right) + \cos\left(\left[\omega_1 + \omega_2\right]t + \phi_1 + \phi_2\right)\right]
\end{aligned}
,
$$

which contains the frequencies

$$0,\ 2\omega_1,\ 2\omega_2,\ \omega_1 + \omega_2, \text{ and } \omega_1 - \omega_2,$$

along with $\omega_1$ and $\omega_2$. For the contributions to $I_{sc}$ from the cubic term, we have the following expression:

$$
\begin{aligned}
&E_{inc}{}^{3} = \left[E_1 \cos\left(\omega_1 t + \phi_1\right) + E_2 \cos\left(\omega_2 t + \phi_2\right)\right]^3 \\
&= E_1{}^{3} \cos\left(\omega_1 t + \phi_1\right)^3 + 3E_1{}^{2} E_2 \cos^2\left(\omega_1 t + \phi_1\right)\cos\left(\omega_2 t + \phi_2\right) \\
&+ 3E_1 E_2{}^{2} \cos\left(\omega_1 t + \phi_1\right)\cos^2\left(\omega_2 t + \phi_2\right) + E_2{}^{3}\cos^3\left(\omega_2 t + \phi_2\right) \\
&= \frac{1}{4}E_1{}^{3}\left[\cos\left(3\omega_1 t + 3\phi_1\right) + 3\cos\left(\omega_1 t + \phi_1\right)\right] \\
&+ \frac{1}{4}E_2{}^{3}\left[\cos\left(3\omega_2 t + 3\phi_2\right) + 3\cos\left(\omega_2 t + \phi_2\right)\right] \\
&+ \frac{3}{2}E_1{}^{2} E_2\left[1 + \cos\left(2\omega_1 t + 2\phi_1\right)\right]\cos\left(\omega_2 t + \phi_2\right) \\
&+ \frac{3}{2}E_1 E_2{}^{2}\cos\left(\omega_1 t + \phi_1\right)\left[1 + \cos\left(2\omega_2 t + 2\phi_2\right)\right] \\
&= \frac{1}{4}E_1{}^{3}\cos\left(3\omega_1 t + 3\phi_1\right) + \left[\frac{3}{4}E_1{}^{3} + \frac{3}{2}E_1{}^{2}E_2\right]\cos\left(\omega_1 t + \phi_1\right) \\
&+ \frac{1}{4}E_2{}^{3}\cos\left(3\omega_2 t + 3\phi_2\right) + \left[\frac{3}{4}E_2{}^{3} + \frac{3}{2}E_1 E_2{}^{2}\right]\cos\left(\omega_2 t + \phi_2\right) \\
&+ \frac{3}{4}E_1{}^{2}E_2\left[\cos\left(\left[2\omega_1 + \omega_2\right]t + 2\phi_1 + \phi_2\right) + \cos\left(\left[2\omega_1 - \omega_2\right]t + 2\phi_1 - \phi_2\right)\right] \\
&+ \frac{3}{4}E_1 E_2{}^{2}\left[\cos\left(\left[2\omega_2 + \omega_1\right]t + 2\phi_2 + \phi_1\right) + \cos\left(\left[2\omega_2 - \omega_1\right]t + 2\phi_2 - \phi_1\right)\right]
\end{aligned}
,
$$

which contains the frequencies

$$3\omega_1, 3\omega_2, \omega_1, \omega_2, 2\omega_1 + \omega_2, 2\omega_1 - \omega_2, 2\omega_2 + \omega_1, \text{ and } 2\omega_2 - \omega_1 .$$

There are two things to note about this result:

(1) the responses at the illuminating radar frequencies $\omega_1$ and $\omega_2$ are no longer linear, which makes the nominal cross-sections at these frequencies field dependent, and

(2) Unlike the frequencies generated by the squared term in $I_{sc}$, the fact that $\omega_1 \approx \omega_2$ implies that $2\omega_1 - \omega_2 \approx \omega_1$ and $2\omega_2 - \omega_1 \approx \omega_2$; i.e., these frequencies lie within the antenna

bandwidth. In mixer theory, these frequencies are referred to as intermodulation products (*5*).

Let us consider only the in-band signals returning to the antenna. Assuming the antenna gain and impedance are the same for frequencies $\omega_1$ and $\omega_2$, the scattered fields at the various radiated frequencies take the form

$$E_{sc}\left(\omega_1\right)=\frac{1}{R}\left(\sigma^{1/2}E_1+\beta\left[\frac{3}{4}E_1{}^3+\frac{3}{2}E_1{}^2E_2\right]\right)$$
$$=\frac{(ZG)^{1/2}}{(4\pi)^{1/2}R^2}P_1^{1/2}\left(\sigma^{1/2}+\frac{3}{4}\beta\frac{ZG}{4\pi R^2}\left[P_1+2P_2\right]\right)$$
$$E_{sc}\left(\omega_2\right)=\frac{1}{R}\left(\sigma^{1/2}E_2+\beta\left[\frac{3}{4}E_2{}^3+\frac{3}{2}E_1E_2{}^2\right]\right)$$
$$=\frac{(ZG)^{1/2}}{(4\pi)^{1/2}R^2}P_2{}^{1/2}\left(\sigma^{1/2}+\frac{3}{4}\beta\frac{ZG}{4\pi R^2}\left[P_2+2P_1\right]\right)$$
$$E_{sc}\left(2\omega_1-\omega_2\right)=\frac{1}{R}\left(\frac{3}{4}\beta E_1{}^2E_2\right)=\frac{3}{4}\beta\frac{(ZG)^{3/2}}{(4\pi)^{3/2}R^4}P_1P_2{}^{1/2}$$
$$E_{sc}\left(2\omega_2-\omega_1\right)=\frac{1}{R}\left(\frac{3}{4}\beta E_1E_2{}^2\right)=\frac{3}{4}\beta\frac{(ZG)^{3/2}}{(4\pi)^{3/2}R^4}P_2P_1{}^{1/2} \quad . \tag{1}$$

Then the fields arriving at the intermod frequencies are

$$E_{rec}\left(2\omega_1-\omega_2\right)=\left(\frac{G}{4\pi}\right)^{1/2}\cdot E_{sc}\left(2\omega_1-\omega_2\right)=\frac{3}{4}\beta\frac{Z^{3/2}G^2}{(4\pi)^2R^4}P_1P_2{}^{1/2},$$
$$E_{rec}\left(2\omega_2-\omega_1\right)=\left(\frac{G}{4\pi}\right)^{1/2}\cdot E_{sc}\left(2\omega_2-\omega_1\right)=\frac{3}{4}\beta\frac{Z^{3/2}G^2}{(4\pi)^2R^4}P_2{}^2P_1$$

and the powers received at the intermod frequencies are

$$P_{rec}\left(2\omega_1-\omega_2\right)=Z^{-1}E_{sc}\left(2\omega_1-\omega_2\right)^2A=\frac{9}{16}\beta^2\frac{Z^2G^4A}{(4\pi)^4R^8}P_1{}^2P_2$$
$$=\frac{9}{16}\beta^2\frac{Z^2G^5\lambda^2}{(4\pi)^5R^8}P_1{}^2P_2 \quad .$$
$$P_{rec}\left(2\omega_2-\omega_1\right)=Z^{-1}E_{sc}\left(2\omega_2-\omega_1\right)^2A=\frac{9}{16}\beta^2\frac{Z^2G^4A}{(4\pi)^4R^8}P_2{}^2P_1$$
$$=\frac{9}{16}\beta^2\frac{Z^2G^5\lambda^2}{(4\pi)^5R^8}P_2{}^2P_1$$

Looking past the complexity of these expressions, let us focus on the *R*–dependence of the intermod terms (*5*). Equation 1 reveals that $P_{rec}\,(2\omega_1-\omega_2)$ and $P_{rec}\,(2\omega_2-\omega_1)$ both fall off as $1/R^8$, rather than the traditional $1/R^4$ predicted by the radar equation. Note that this drastic attenuation of receiver power at the new frequencies will be somewhat mitigated by the increased antenna gain coefficient $G^5$.

At first glance, it appears that we have identified a near-field component of the radiation from the scatterer, since there is clearly no power at infinity from a field that falls off faster than $1/R$. To counter this assertion, let us consider the bistatic case. Let the transmit antenna have a gain $G_t$ and be a distance $R_t$ from the target, while the receive antenna has a gain $G_r$ and is at a distance $R_r$ from the target. Assume that both $R_t$ and $R_r$ are large compared to wavelength and target dimensions. Then the same derivation as above leads to the following expression for the intermod fields at the scatterer:

$$E_{sc}\left(2\omega_1-\omega_2\right)=\frac{1}{R_r}\left(\frac{3}{4}\beta E_1{}^2E_2\right)=\frac{1}{R_r}\left(\frac{3}{4}\beta\frac{\left(Z_tG_t\right)^{3/2}}{(4\pi)^{3/2}R_t{}^3}P_1P_2{}^{1/2}\right).$$

$$E_{sc}\left(2\omega_2-\omega_1\right)=\frac{1}{R_r}\left(\frac{3}{4}\beta E_1E_2{}^2\right)=\frac{1}{R_r}\left(\frac{3}{4}\beta\frac{\left(Z_tG_t\right)^{3/2}}{(4\pi)^{3/2}R_t{}^3}P_2P_1{}^{1/2}\right)$$

Then the field detected at the receiver is of the form

$$E_{rec}\left(2\omega_1-\omega_2\right)=\left(\frac{G_r}{4\pi}\right)^{1/2}\cdot E_{sc}\left(2\omega_1-\omega_2\right)=\frac{3}{4}\beta\frac{Z_t{}^{3/2}G_r{}^{1/2}G_t{}^{3/2}}{(4\pi)^2R_rR_t{}^3}P_1P_2{}^{1/2}.$$

$$E_{rec}\left(2\omega_2-\omega_1\right)=\left(\frac{G_r}{4\pi}\right)^{1/2}\cdot E_{sc}\left(2\omega_2-\omega_1\right)=\frac{3}{4}\beta\frac{Z_t{}^{3/2}G_r{}^{1/2}G_t{}^{3/2}}{(4\pi)^2R_rR_t{}^3}P_2{}^2P_1$$

Now the intermod products have a factor of $\frac{G_t{}^{3/2}G_r{}^{1/2}}{R_t{}^3R_r}$, and the powers received at the intermod frequencies go as $\frac{G_t{}^3G_r}{R_t{}^6R_r{}^2}$. Clearly, the initial trip from transmitter to target satisfies the far-field condition by construction, while the $1/R_r^2$ behavior arising from the trip from target to receive antenna shows *prima faciae* that it is also satisfied on this leg of the trip, as well. There will, in fact, be near-field contributions to the radiation from the scatterer, but they contain higher powers of $1/R_r^2$ and, hence, are even more attenuated by distance. Note that in the bistatic case, the coefficient β may depend on the angle between the receiver and transmitter boresights.

The nonlinear radar cross-section becomes very small when it involves higher order terms. For example, a 1 MW source would barely have 300 m range.

## 4. Multipath Effects

In situations where the radar and the target are both relatively close to the ground, multipath effects can further exacerbate this range-dependent attenuation. Although in rare cases multipath may actually enhance a target return, the most likely effect at low grazing angles is destructive interference caused by the ground reflection. Vertical polarization can provide some advantages due to pseudo-Brewster angle effects, but the Brewster angle ranges from about 13° (moist soils) to about 30° (arid soils), while the grazing angle from a 2-m high sensor to a point 100 m away is 1.1°. At these shallow angles, the behavior of both horizontally and vertically polarized waves is essentially the same, with an inversion of sign upon reflection, and with very little loss in amplitude.

In general, the effect of multipath cancellation is to add another $1/R^2$ to the power dependence each way. To demonstrate where this dependence comes from, consider a radar located near the ground illuminating a target near the ground. To simplify the model, consider the geometry shown in figure 1. An antenna at a height *h* off the ground illuminates a point on a target a horizontal distance *R*, where the point is a height *s* off the ground. The signal from the antenna is taken to be the sum of a direct propagation from antenna to target and an indirect propagation involving a bounce off the ground at a horizontal distance αR from the target, with α < 1. Assuming that the reflection causes a complete phase reversal with no loss of amplitue (again, this is a reasonable assumption for shallow grazing angles, or purely horizontal polarization), the signal at the target is of the form:

$$E = E_0(R)\left(\cos\frac{\omega}{c}D - \cos\frac{\omega}{c}\left(D_1 + D_2\right)\right),$$

where $E_0(R) \propto \frac{1}{R}$ as before, and

$$\begin{aligned} D_1 &= \sqrt{h^2 + (1-\alpha)^2 R^2} \\ D_2 &= \sqrt{s^2 + \alpha^2 R^2} \\ D &= \sqrt{(h-s)^2 + R^2} \end{aligned} .$$

At distances $R >> h, s$ we find that

$$D_1 + D_2 - D \approx \frac{1}{2R}\left[h^2\left(\frac{\alpha}{1-\alpha}\right) + s^2\left(\frac{1-\alpha}{\alpha}\right) - 2hs\right]$$

and

$$E \approx \frac{1}{2} E_0 \cos\frac{\omega}{2c}\left(D_1 + D_2 + D\right)\cos\frac{\omega}{2c}\left(D_1 + D_2 - D\right)$$
$$\propto \frac{1}{R^2}\cos\frac{\omega}{c}D$$

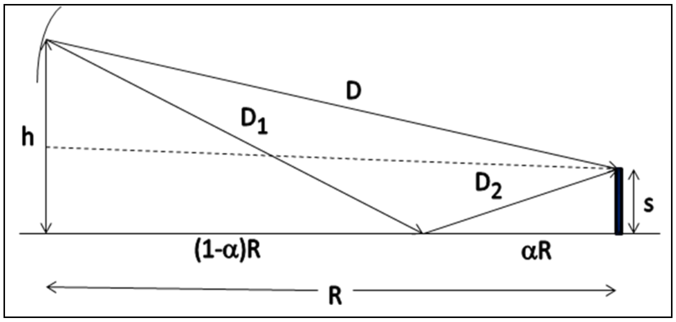


Figure 1. Multipath geometry.

Clearly, this change in amplitude at the target corresponds to a $1/R^4$ change in the power from antenna to target, and a $1/R^8$change for the round trip, even for a linear target. In figure 2, we plot the amplitude, in dB, versus the range to the target for a radar located 2 m off the ground illuminating a target 0.25 m off the ground. The 6 dB peaks in the signals for the 2000 and 3000 MHz curves is due to constructive interference, but farther out in range, all of the curves approach a 20 dB/decade fall-off in amplitude, thus demonstrating the $1/R^2$ rule. Consider the region between 15 and 20 m in the example of figure 2. Here the illumination near 1000 MHz is definitely influenced by an additional $1/R^2$ term and, hence, a $1/R^8$ falloff is expected. Combining this with the effect of the nonlinear target gives a total falloff of $1/R^{16}$. This would mean a change of range from 50 to 100 m would result in a loss of signal of $2^{16}$ or 48 dB per octave or 160/dB per decade.

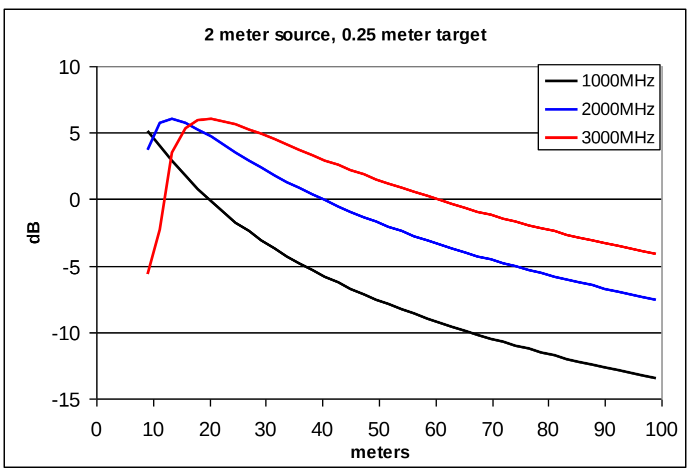


Figure 2. Range dependence of signal in a multipath environment.

Multipath is of particular importance when the target height above ground can vary quickly, as is the case when insects are tagged with non-linear transponders (*1, 2*). In flight, the insect can appear as a free-space target, but as it approaches, or crawls along the ground, its effective cross-section is seriously reduced, as can be seen in figure 3. Here we examine the effects of varying the target height at a fixed range of 100 m from the radar. Note that there are situations where additional nulls can occur (e.g., 3000 MHz at 2.5 m height, 2000 MHz at 3.75 m height), and this happens more frequently at smaller wavelengths and may cause a fast variation in target signature as the target height varies.

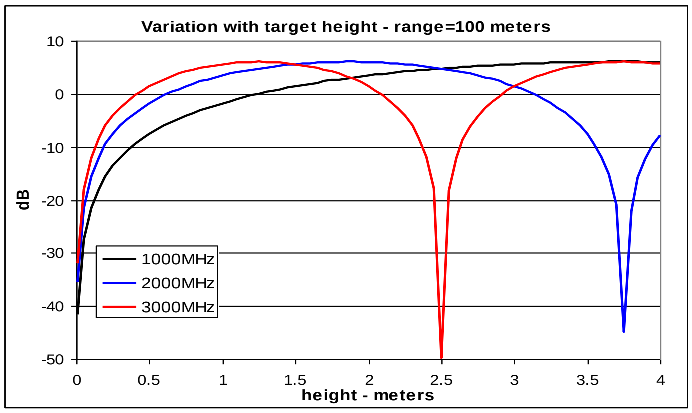


Figure 3. Change in target strength versus height of target for a target range of 100 m.

## 5. Summary

We emphasize here that we are not advocating any modification of the radar equation, itself. Rather, in this paper we are simply asserting that in nonlinear radar, the radar cross-section (RCS) of a target is no longer independent of distance from the source, and has to be redefined for each order of nonlinearity in the target response.

# 6. References


1. O'Neal, M. E.; Landis, D. A.; Rothwell, E.; Kempel, L.; Reinhard, D. Tracking Insects with Harmonic Radar: a Case Study. *American Entomologist*, Winter 2004, pp 212–218.

2. Charvat, G. L.; Rothwell, E. J.; Kempel, L. C. Harmonic Radar Tag Measurement and Characterization. *IEEE Antennas and Propagation Society International Symposium* **June 2003**, *2*, 696–699.

3. Jackson, J. D. *Classical Electrodynamics*; 2nd ed. John Wiley & Son: New York, 1975, Sec. 9.6.

4. Hovanessian, S. A. *Radar System Design and Analysis*; Artech House Inc.: Dedham MA, 1984), Ch. 1.

5. Maas, S. *Microwave Mixers*; Artech House Inc.: Norwood MA, 1993, Sec. 5.2.

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